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(54) Title: ESTIMATION OF DOPPLER SHIFT IN SEVERAL STEPS

(57) Abstract: Method and system for carrier recovery and estimation of Doppler shift from a signal source that is moving relative to a signal receiver. A pure carrier preamble for the received signal is processed through each of two stages of a linear predictor to obtain a successively more accurate estimation of a Doppler frequency offset for the carrier. The received signal is downconverted by each stage estimation of the Doppler frequency offset, and the downconverted signal is processed through a decision feedback phase locked loop to provide a signal in which substantially all of the Doppler offset and/or phase angle are identified and removed. The system has low complexity, is fast, and is accurate to within an estimated few tens of Hertz and will work with signals having relatively low signal-to-noise ratios. The invention is useful for receipt of signals from satellites in low earth orbits (LEOs) and other non-geosynchronous orbits, and wherever a transmitter and receiver are moving relative to each other.

ESTIMATION OF DOPPLER SHIFT IN SEVERAL STEPS

Field of the Invention

This invention relates to digital filtering of multiple bits input signals and computation of correlation or comvolution values.

Background of the Invention

In formation of correlation values from digital signals, the number of multiplications required to form a single correlation values can be as large as the number P of time slots or "chips" used to define the signal, if a conventional approach is used for computation of the correlation values. In a code division multiple access (CDMA) approach to communications, the communication symbols are multiplied or "spread" using a fixed pattern of reference symbols, referred to as pseudo-random (PN) code, and transmitted. At the receiver end, an optimal solution for detecting the received symbols, and identifying the particular code used for the spreading, employs a maximum likelihood (ML) decision rule. The ML solution for coded symbols is a matched filter or correlator, with coefficients given by the same PN code, a process known as "despreading". Analysis of a correlation curve for the received signal and determination of the time shift and amplitude of its peak value allows two questions to be answered: (1) Is the reference pattern used to form the correlation curve likely to be the same as the spreading pattern used at the transmitter; and (2) If the answer to the first question is "yes", what is the most likely time shift associated with the received signal.

A CDMA communication system receiver requires use of at least 3·N matched filter values, where N is the number of rake fingers used for the communications. The matched filter coefficients used for despreading account for the largest fraction of the gate count in a CDMA chip and have an associated built-in time delay for the substantial signal processing that is required. For example, in use of Code Division Multiple Access (CDMA) for signal recognition in a Global Positioning System (GPS), the number of chips used is P = 1023, which requires about 1023 adders, arranged in about ten layers or more, to compute a single correlation value, and this computation is repeated for each of the approximately 1023 time shifts in order to compute a

set of values for a full correlation curve. The time delay for (a minimum of) ten processing layers can be many tens of nanoseconds.

A second category where PN codes are used is in frame synchronization in digital communications. The transmitted information symbols are divided and packed into frames that also carry header and trailer information (overhead), including a preamble for synchronization of the remainder of an asynchronously transmitted frame. A PN code or "unique word" (UW) is included to assist in detecting the beginning point of the payload within a frame. The receiver searches for the UW within a frame, and an optimal solution is an ML synchronizer. Again, this requires use of a matched filter or correlator, with coefficients given by the UW used in the frame. Again, the matched filter coefficients used for pattern recognition account for the largest fraction of the gate count in frame synchronization.

What is needed is an approach that reduces the number of arithmetic operations used in formation of a correlation value used in CDMA communications, in frame synchronization and in any other signal processing procedure requiring formation of one or more correlation values. Preferably, the approach should have reduced complexity and should allow use of a much smaller gate count and have a smaller associated time delay for computation of each correlation value. Preferably, the approach should be flexible enough to accommodate a signal of any magnitude, using any signal representation format (binary, quaternary, octal, binary coded decimal, hexadecimal, etc.) and using any signal over-sampling rate.

Summary of the Invention.

These needs are met by the invention, which reduces the number of addition operations from M·N·R to a maximum of M·N, where M is the maximum number of orders of magnitude needed to represent a signal magnitude (e.g., M=1 for up to 10, M=2 for up to 100, etc. in decimal base), N is the number of binary digits used to represent a single "order of magnitude" (e.g., N=3 for octal, N=4 for a decimal or hexadecimal order of magnitude), and R is the over-sampling rate used for signal reception (e.g., R = 2-10, or higher). The invention computes differences between two consecutive correlation values, rather than computing each correlation value separately, and this allows reduction, by a multiplicative factor of at least R, depending upon

the structure of the reference or coding word, in the number of arithmetic operations for computation of a correlation value. The number of gates required is reduced by the same factor, and the associated time delay for computation of a correlation value is reduced by at least $\Delta t(\text{add})\cdot\log_2(R)$, where $\Delta t(\text{add})$ is the time required to perform an addition of two values. Optionally, a time shift point at which the correlation curve reaches a maximum is automatically identified by this approach, and the corresponding (maximum) correlation value is quickly recovered after this time shift point has been identified.

Brief Description of the Drawings

Figure 1 is a graphical view of a representative correlation curve for a received signal.

Figure 2 illustrates a gate array used to compute correlation values according to a conventional approach.

Figure 3 illustrates overlapping arrays of over-sampled PN Code values (± 1) used in an example that explains use of the invention.

Figure 4 illustrates a gate array used to compute correlation values according to the invention, corresponding to the conventional approach in Figure 2.

Detailed Description of the Invention

Figure 1 is a graph of an idealized autocorrelation function that might be formed from a received digital signal s(t;rec), which is generally expressible as a multi-bit numerical value. Each time point t_n corresponds to a time shift for a convolution value $C(t_n)$, computed as

$$C(t_n) = \sum_{k=1}^{K} s(t_k;\text{rec})\cdot s(t_n - t_k;\text{ref}), \qquad (1)$$

where s(t;ref) is the reference signal used at the receiver to implement pattern recognition for the received signal and $\{t_k\}$ is a sequence of time points (not necessarily equidistantly spaced) used to compute a correlation value for the received signal. The time shift $\Delta t(\text{max})$ corresponding to the position where the maximum amplitude of the correlation curve in Figure 1 may be, but need not be, precisely equal to one of the time shift values t_n used to compute the convolution values.

Use of the invention is first illustrated by an example. A PN code signal has an amplitude s(t;ref) = 789, which is represented in hexadecimal format as an ordered sequence 0111/1000/1001. The received signal is over-sampled at a selected rate R, for example, R = 4, so that each of the 12 signal values representing the received signal s(t;rec) is appears four consecutive times. A "modified hexadecimal format", in which the binary value "0" is replaced by the numerical value "-1", is adopted here to more easily illustrate operation of the invention. A modified form of the invention will also work with the conventional hexadecimal format representation. Adopting the modified hexadecimal format, the over-sampled expression for PN Code amplitude s(t;ref) = 789 becomes

$$\{s(t;ref)\} = -1\ -1\ -1\ -1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ 1\ -1\ -1\ -1\ -1$$
$$-1\ -1\ -1\ -1\ -1\ -1\ -1\ -1\ 1\ 1\ 1\ 1\ -1\ -1\ -1\ -1\ -1\ -1\ -1\ -1\ 1\ 1\ 1\ 1. \qquad (2)$$

This number has 48 = 3·4·4 binary digits in its representation, including factors corresponding to an order of magnitude number, M = 3, a hexadecimal (or decimal) representation number, N = 4, and an over-sampling rate number, R = 4, in this example. The time shifts t_n are equidistant so that $t_{n+1} - t_n = \Delta t =$ constant..

In a conventional approach to computation of the correlation values using this PN Code, a sequence of time shifts t_n is applied to the over-sampled reference signal values $s(t_k;ref)$, and the resulting signal values are multiplied by the received signal values, as indicated in Eq. (1). For any selected value of the time shift t_n, computation of the corresponding correlation value requires use of 48 signal value multipliers and use of 47 (or 48) adders, which may be arranged six layers containing 24, 12, 6, 3, 2 and 1 two-input adders as shown in Figure 2. The associated time delay Δt(conv) for conventionally processing a received signal to provide a single correlation value is six times the time delay Δt(add) associated with addition of two values at a single adder. More generally, the associated time delay is approximately $\{\log_2[M\cdot N\cdot R]\}_+$, where $\{P\}_+$ is the smallest integer that is greater than or equal to the real number P; this is related to, but not the same as, the integer part, $[P]_+$, of the real number P. For this example, the associated time delay is $\Delta t(add)\cdot\{\log_2[3\cdot 4\cdot 4]\}_+ =$ $\Delta t(add)\cdot\{5.585\}_+ = 6\cdot\Delta t(add)$.

For the example chosen, the convolution values $C(t_n)$ defined in Eq. (1) become

$$C(n\cdot\Delta t) = \sum_{k=1}^{K} s(k\cdot\Delta t;rec)\cdot PN((n-k)\cdot\Delta t), \qquad (3)$$

where the sequence of PN code values is set forth in Eq. (2). The difference of two consecutive convolution values is verified to be

$$\begin{aligned}\Delta C(t_{n+1}) &= C(t_{n+1}) - C(t_n)\\ &= \sum_{k=1}^{47} s(k\cdot\Delta t;rec)\cdot\{PN((n+1-k)\cdot\Delta t) - PN((n-k)\cdot\Delta t)\}\\ &\quad + s(48\cdot\Delta t;rec)\cdot PN((n+1-48)\cdot\Delta t) - s(0\cdot\Delta t;rec)\cdot PN(n\cdot\Delta t)\\ &= \sum_{k=0}^{48} s(k\cdot\Delta t;rec)\cdot w(n;k), \qquad (4)\end{aligned}$$

$$\begin{aligned}w(n;k) &= 0 \quad (n\text{-}k \text{ not divisible by } 4)\\ &= PN((n+1-48)\cdot\Delta t) \quad (k = 48)\\ &= -PN(n\cdot\Delta t) \quad (k = 0)\\ &= 2 \quad (n\text{-}k = 4\cdot u;\ PN(4\cdot u) = -PN(4\cdot u + 1) = -1;\ k \neq 0, 48)\\ &= 0 \quad (n\text{-}k = 4\cdot u;\ PN(4\cdot u) = PN(4\cdot u + 1)\ )\\ &= -2 \quad (n\text{-}k = 4\cdot u;\ PN(4\cdot u) = -PN(4\cdot u + 1) = +1), \qquad (5)\end{aligned}$$

where $u = 0, \pm1, \pm2, \pm3, \ldots$. One verifies that the convolution value difference, $C((n+1)\cdot\Delta t) - C(n\cdot\Delta t)$, includes at most $13 = 3\cdot4 + 1$ terms, corresponding to the integer values of k in the convolution sum in Eq. (3) for which the integer difference, n-k, is divisible by 4 (n-k = 0, 1, 2, ..., 48). In the particular example with a PN code of 789, of the indices for which n-k is divisible by 4, a sequence of 13 consecutive values of the coefficients w(n;k) becomes

$$\{w(n;k)\} = \{-1, 2, 0, 0, -2, 0, 0, 0, 2, -2, 0, 2, 1\} \qquad (6)$$

with the first and thirteenth values w(n;k) (= +1 or -1) being determined by individual PN Code values, rather than Code differences.

This example has used the particular choices

M = maximum order of magnitude of reference signal = 3,

N = number of binary digits used for each order of magnitude = 4,

R = over-sampling rate = 4.

Figure 3 illustrates the 48 values of the PN Code represented in Eq. (2), for the shifted time values s(t;ref), s(t+Δt;ref) and s(t+2Δt;ref). Each of these three sequences of 48 numbers is multiplied by corresponding members of a digital signal sequence s(t;rec), as indicated in Eq. (1). Because of the over-sampling rate (R=4) adopted in this example, every fourth value of the sequence in Figure 3 representing s(t;ref) is a potential "boundary point", where the PN Code sequence members s(t;ref) and s(t+Δt;ref) can have different signs (e.g., +1/-1 or -1/+1). Similarly, every fourth value of the sequence representing s(t+Δt;ref) is a potential "boundary point", where s(t+Δt;ref) and s(t+2Δt;ref) have different signs. Where two time-shifted sequences. such as s(t;ref) and s(t+Δt;ref) have the same sign (e.g., +1/+1 or -1/-1) at such a boundary point, this boundary point will not contribute a non-zero value to a convolution value difference $C(t_n) - C(t_{n-1})$.

Figure 4 illustrates an array 41 of coefficient multipliers and signal summers that can be used to form a convolution value difference for the example considered here, with M = 3, N = 4 and R = 4 but with an arbitrary PN code. The top level of the array has a maximum of M·N-1 = 11 coefficient multiplier modules 43-k, each with an associated coefficient w(n;k) for k = 1, 2, ..., 11 that has a value +2, 0, or -2, depending upon the particular PN Code used, plus one signal differencer that receives the signals s(48·Δt;rec) and s(0;rec) and forms and issues a weighted difference

$$\Delta s(n) = s(48\cdot\Delta t;rec)\cdot PN((n+1-48)\cdot\Delta t) - s(0;rec)\cdot PN(n\cdot\Delta t). \qquad (7)$$

The output signal for the coefficient multiplier module 43-k is

$$O(43\text{-}k) = w(n;k)\cdot s(4\cdot k\cdot\Delta t;rec). \qquad (8)$$

For the PN Code 789 used in the preceding example,

$$w(n;1) = w(n;8) = w(n;11) = +2,$$
$$w(n;4) = w(n;10) = -2.$$
$$w(n;2) = w(n;3) = w(n;5) = w(n;6) = w(n;7) = w(n;9) = 0,$$

as can be verified from Eq. (6). The second array level from the top has six two-input signal adders 45-j (j = 1, 2, ..., 6) producing six output sum signals. The third array level from the top has three two-input signal adders 47-h (h = 1, 2, 3), producing three output sum signals. The fourth array level from the top has one two-input adder 49; and the fifth array level from the top has one two-input signal adder 51, whose output signal is the convolution value

difference $\Delta C(t_n)$ for some index value n. The array 41 in Figure 4 has M·N-1 = 11 signal coefficient multipliers at the top level and four levels with an additional 12 signal adders. The maximum associated time delay is approximately 5·Δt(add). The number of gates required is a maximum of 23 (18 for the PN Code 789 chosen for this example), as compared with 47 for the conventional approach for convolution.

More generally, the convolution value set forth in Eq. (4) can be expressed as

$$\Delta C(t_{n+1}) = C(t_{n+1}) - C(t_n)$$

$$= \sum_{k=0}^{M\cdot N\cdot R} s(k\cdot\Delta t;rec)\cdot\{PN((n+1-k)\cdot\Delta t) - PN((n-k)\cdot\Delta t)\}$$

$$= \sum_{k=0,}^{M\cdot N\cdot R} s(k\cdot\Delta t;rec)\cdot W(n;k), \tag{9}$$

$$\begin{aligned} W(n;k) &= 0 \quad (\text{n-k not divisible by R}) \\ &= PN((n+1-M\cdot N\cdot R)\cdot\Delta t) \quad (k = M\cdot N\cdot R) \\ &= -PN(n\cdot\Delta t) \quad (k = 0) \\ &= 2 \quad (n-k = R\cdot u;\ PN(R\cdot u) = -PN(R\cdot u + 1) = -1) \\ &= 0 \quad (n-k = R\cdot u;\ PN(R\cdot u) = PN(R\cdot u + 1)) \\ &= -2 \quad (n-k = R\cdot u;\ PN(R\cdot u) = -PN(R\cdot u + 1) = +1), \end{aligned} \tag{10}$$

and the maximum number of terms to be added to compute the convolution difference is M·N + 1. Multiplication by a factor of 2 is implemented using a one-bit shift of a binary-expressed value.

The innovative procedure begins with computation or provision of a first convolution value

$$C(t_1) = \sum_{k=0}^{M\cdot N\cdot R} s(t_k;rec)\cdot s(t_1 - t_k;ref). \tag{11}$$

This first convolution value may be known from other considerations (e.g., $C(t_1) = 0$) and may not need to be computed. For $n \geq 1$, only the convolution differences, $\Delta C(t_{n+1}) = C(t_{n+1}) - C(t_n)$, as set forth in Eq. (9), are computed. The total number of additions (and multiplications by 2) required here is at most M·N and may be less, depending upon the values of the coefficients

W(n;k) as set forth in Eq. (10). One can verify that the number of layers of two-input adders needed to accomplish these M·N value additions is at most $\{\log_2[M\cdot N\}_+$. If K-input adders (K≥2) are substituted for the two-input adders, the number of layers of K-input adders needed to accomplish these M·N value additions is reduced to at most $\{\log_K[M\cdot N\}_+$. Each factor-of-two multiplication required by the coefficients W(n;k) in Eq. (10) can be implemented by shifting left by one binary digit in a shift register. If a particular convolution value, say $C(t_m)$, is needed, this value is computed using the relation

$$C(t_m) = C(t_1) + \sum_{n=1}^{m} \{C(t_n) - C(t_{n-1})\}. \qquad (12)$$

For the particular example discussed in the preceding, Figure 4 illustrates a suitable array to compute the convolution value differences. This array requires at most $\{\log_2[M\cdot N\}_+ = 4$ layers, as opposed to $\{\log_2[M\cdot N\cdot R]\}_+, = 6$ for the conventional approach, and requires at most M·N = 12 adders, as opposed to M·N·R = 48 for the conventional approach.

The integer M depends upon the "order of magnitude" chosen for the PN code. The preceding discussion relies upon use of a hexadecimal format, in which N=4 consecutive binary digits are used to express each order of magnitude value, such as a decimal. Use of a decimal format would also require N=4, because use of at least four binary digits is required to express the values 8 and 9 in a binary system. One can also express an order of magnitude value in octal format, using N=3 consecutive binary digits. In an octal format, the largest order of magnitude value becomes 7 (expressed as 111 in an octal format), not the maximum order of magnitude value 9 of a decimal format or the value F (equivalent to 15) of a hexadecimal format.

The innovative method disclosed here uses single unit differences $\Delta C(t_{n+1}) = C(t_{n+1}) - C(t_n)$, as set forth in Eq. (9), to quickly compute the convolution values themselves, as set forth in Eq. (12). Double unit differences

$$\begin{aligned}\Delta_2 C(t_{n+2}) &= C(t_{n+2}) - C(t_n)\\ &= \Delta C(t_{n+2}) + \Delta C(t_{n+1})\end{aligned} \qquad (13)$$

can be computed in a similar manner, if desired, using a sum of single unit differences, as computed in the preceding. More generally, one can compute p-unit differences

$$\Delta_p C(t_{n+p}) = C(t_{n+2}) - C(t_n)$$

$$= \sum_{k=1}^{p} \Delta C(t_{n+k}) \qquad (14)$$

in a similar manner, if desired.

The innovative approach disclosed here is also useful in estimating the time shift value $\Delta t(\max)$ that corresponds to a local maximum value of the convolution value $C(t_n)$, where $\Delta t(\max)$ need not coincide with one of the discrete time shift values t_n but may lie between two discrete time shift values. Assume that t_{m-1}, t_m, and t_{m+1}, are three consecutive time shift points for the correlation curve in Figure 1, and that the time shift value $\Delta t(\max)$ lies between t_{m-1} and t_{m+1}. This situation assumes that the consecutive convolution values $C(t_{m-1})$, $C(t_m)$ and $C(t_{m+1})$ satisfy the constraint

$$C(t_m) \geq \max\{C(t_{m-1}), C(t_{m+1})\}. \qquad (15)$$

This constraint may be expressed in an equivalent form for convolution signal differences as a pair of constraints

$$\Delta C(t_m) \geq 0\ , \qquad (16A)$$

$$\Delta C(t_{m+1}) \leq 0. \qquad (16B)$$

The value $\tau(\max)$ can be computed as follows. A second degree (or higher degree) polynomial fit to the correlation values at the time shift points t_{m-1}, t_m, and t_{m+1}, becomes

$$P(\tau;2) = c0 + c1\cdot\tau + c2\cdot\tau^2 \quad (t_{m-1} \leq \tau \leq t_{m+1}), \qquad (17)$$

$$c2 = \{[C(t_{m-1}) - C(t_{m+1})]\cdot(t_m - t_{m+1}) + [C(t_m) - C(t_{m+1})]\cdot(t_{m+1} - t_{m-1})\}/(\det), \qquad (18)$$

$$c1 = \{[C(t_{m-1}) - C(t_{m+1})]\cdot(t_{m+1}{}^2 - t_m{}^2) + [C(t_m) - C(t_{m+1})]\cdot(t_{m-1}{}^2 - t_{m+1}{}^2)\}/(\det), \qquad (19)$$

$$c0 = C(t_m) - c1\cdot(t_m) - c2\cdot(t_m)^2, \qquad (20)$$

$$\det = \{(t_{m-1} - t_m)\cdot(t_m - t_{m+1})\cdot(t_{m-1} - t_{m+1})\}^{-1}. \qquad (21)$$

The time shift for maximum amplitude is given by

$$\tau(\max) = -c1/c2, \qquad (22)$$

and the time shift value $\tau(\max)$ depends only upon the two correlation signal differences, $\Delta C(t_m) = C(t_m) - C(\Delta t_{m-1}) - C(\Delta t_{m+1})$ and $\Delta C(t_{m+1}) = C(\Delta t_{m+1}) - C(\Delta t_m)$, which have already been computed according to the procedure set forth in the preceding.

More generally, the second degree polynomial $P(\tau;2)$ can be replaced by an nth degree polynomial $P(\tau;n)$ $(n\geq 2)$, with appropriate constraints imposed on the polynomial values at three or more of the discrete time points $\{t_k\}$, and the time shift value $\tau(\max)$ that maximizes the polynomial $P(\tau;n)$ within the interval $t_{m-1} \leq \tau \leq t_{m+1}$ can be determined.

In one interpretation, the time shift value $\tau(\max)$ is identified as an estimate of the time shift value corresponding to the maximum value the convolution value $C(t_k)$ would attain, if the convolution were extended to continuous values of the time variable t.

In a second interpretation, one imposes the conditions (16A) and (16B) and determines or computes a time variable value t = tm' for which

$$|t_{m'} - \tau| = \min\{ |t_{m'} - t_{m-1}|, |t_{m'} - t_m|, |t_{m'} - t_{m+1}| \}. \quad (23)$$

The time variable value $t_{m'}$ is then interpreted as the time shift value, drawn from the discrete set $\{t_k\}$, that produces the maximum convolution value.

In a third interpretation, one imposes the conditions (16A) and (16B) and determines or computes a time variable value $t = t_{m''}$ for which

$$C(t_{m''}) = \max\{ C(t_{m-1}), C(t_m), C(t_{m+1}) \}, \quad (24)$$

which will normally be $t_{m''} = t_m$.

The time shift value $t = \tau(\max)$ or $t = t_{m'}$ or $t = t_{m''}$ can be used to determine the most likely time shift in CDMA communications, the most likely point at which the payload (data) begins in frame synchronization, or for any other similar purpose involving use of convolution signals.

What is claimed is:

1. A method for carrier signal recovery in the presence of a Doppler shift, the method comprising:

providing a carrier signal that may contain a Doppler-shifted replica of a desired signal;

processing a first portion of the carrier signal through a first linear predictor to obtain a first contribution to Doppler frequency offset for the received signal;

modifying a second portion of the carrier signal to obtain a first-modified signal from which the first offset contribution is removed;

processing the first-modified signal through a second linear predictor to obtain a second contribution to Doppler frequency offset for the received signal;

modifying the first-modified signal to obtain a second-modified signal from which the second offset contribution is removed; and

processing the second-modified signal through a phase locked loop circuit to estimate at least one of a residual contribution to Doppler-frequency offset and a phase angle for the provided signal.

2. The method of claim 1, further comprising estimating said Doppler frequency offset for said carrier signal as a sum of said first, second and third Doppler frequency offset contributions.

3. The method of claim 1, further comprising providing as said first portion of said carrier signal a carrier signal segment including at most 10 symbols.

4. The method of claim 3, further comprising providing as said second portion of said carrier signal a second carrier signal segment including at most 30 symbols.

5. The method of claim 1, wherein said process of modifying said second portion of said carrier signal to obtain said first-modified signal comprises

multiplying said second portion of said carrier signal by a sinusoidal signal having a frequency equal to said first contribution to said Doppler frequency offset.

6. The method of claim 5, wherein said process of modifying said first-modified signal to obtain said second-modified signal comprises multiplying said first-modified signal by a sinusoidal signal having a frequency equal to said second contribution to said Doppler frequency offset.

7. The method of claim 1, wherein said step of processing said second modified signal through said phase locked loop comprises processing said second modified signal through a decision feedback phase locked loop to estimate at least one of said residual contribution to said Doppler-frequency offset and said phase angle for said provided signal.

8. A method for carrier signal recovery in the presence of a Doppler shift, the method comprising:

providing a first carrier signal s1(t) that may contain a Doppler-shifted replica of a desired signal;

computing a first correlation function

$$S1(N1;K1;T_S) = \sum_{k=1}^{N1} s1(k\cdot T_S;carr)\cdot s1((K1-k)\cdot T_S;carr)^*$$

of the carrier signal with itself, where N1 ($\geq$1) is a first selected sample size, $K1\cdot T_S$ is a first selected time delay and $1/T_S$ is a selected signal sampling rate;

computing a first Doppler frequency offset component

$$f_{D1} = \tan^{-1}\{S1(N1;K1;T_S)\}/(2\pi\cdot K1\cdot T_S);$$

forming a first-modified carrier signal

$$s2(t;carr) = s1(t;carr)\ \exp(-j2\pi\cdot f_{D1}\cdot t);$$

computing a second correlation function

$$S2(N2;K2;T_S) = \sum_{k=1}^{N2} s2(k\cdot T_S;carr)\cdot s2((k-K2)\cdot T_S;carr)^*$$

of the modified carrier signal with itself, where N2 (≥ 1) is a second selected sample size and $K2 \cdot T_S$ is a second selected time delay;

computing a second Doppler frequency offset component

$$f_{D2} = \tan^{-1}\{S2(N2;K2;T_S)\}/(2\pi \cdot K2 \cdot T_S);$$

forming a second-modified carrier signal

$$s3(t;carr) = s2(t;carr)\ \exp(-j2\pi \cdot f_{D2} \cdot t);\ \text{and}$$

processing the second-modified signal through a phase locked loop circuit to estimate at least one of a residual contribution to a Doppler-frequency offset and phase angle for the provided signal.

9. The method of claim 8, further comprising estimating said Doppler frequency offset for said received carrier signal as a sum of said first, second and third Doppler frequency offsets.

10. The method of claim 8, further comprising:

forming a second modified carrier signal

$$s3(t;carr) = s2(t;carr)\ \exp(-j2\pi \cdot f_{D2} \cdot t);\ \text{and}$$

processing the second modified carrier signal by feedback decision phase locked loop analysis to obtain an estimate of at least one of a third Doppler frequency shift and a phase associated with the second modified carrier signal.

11. The method of claim 8, further comprising choosing said integers N1 and N2 to satisfy the relation $N1 \leq N2$.

12. The method of claim 11, further comprising choosing said integers N1 and N2 to be N1 = 60 and N2 = 180.

13. The method of claim 8, further comprising choosing said numbers K1 and K2 to satisfy the relation $D1 \leq D2$.

14. The method of claim 8, further comprising choosing said numbers K1 and K2 to be K1 = 1 and K2 = 20.

15. The method of claim 8, wherein said step of processing said second-modified signal through said phase locked loop circuit further comprises processing said second-modified signal through a decision feedback phase locked loop circuit to estimate at least one of said residual contribution to said Doppler frequency offset and said phase angle for said provided signal.

16. A system for carrier signal recovery in the presence of a Doppler shift, the system comprising:

a signal source that provides a carrier signal that may contain a Doppler-shifted replica of a desired signal; and

a computer that is programmed:

to process a first portion of the carrier signal through a first linear predictor to obtain a first contribution to Doppler frequency offset for the provided signal;

to modify a second portion of the carrier signal to obtain a first-modified signal from which the first offset contribution is removed;

to process the first-modified signal through a second linear predictor to obtain a second contribution to Doppler frequency offset for the provided signal;

to modify the first-modified signal to obtain a second-modified signal from which the second offset contribution is removed; and

to process the second-modified signal through a phase locked loop circuit to estimate at least one of a residual contribution to Doppler-frequency offset and a phase angle for the provided signal.

17. The system of claim 16, wherein said computer is further programmed to estimate said Doppler frequency offset for said carrier signal as a sum of said first, second and third Doppler frequency offset contributions.

18. The system of claim 16, wherein said computer is further programmed to provide as said first portion of said carrier signal a carrier signal segment including at most 10 symbols.

19. The system of claim 18, wherein said computer is further programmed to provide as said second portion of said carrier signal a second carrier signal segment including at most 30 symbols.

20. The system of claim 16, wherein said computer is further programmed to modify said second portion of said carrier signal to obtain said first-modified signal by multiplying said second portion of said carrier signal by a sinusoidal signal having a frequency equal to said first contribution to said Doppler frequency offset.

21. The method of claim 20, wherein said computer is further programmed to modify said first-modified signal to obtain said second-modified signal by multiplying said first-modified signal by a sinusoidal signal having a frequency equal to said second contribution to said Doppler frequency offset.

22. The system of claim 16, wherein said computer is further programmed to process said second modified signal through said phase locked loop by processing said second modified signal through a decision feedback phase locked loop to estimate at least one of said residual contribution to said Doppler-frequency offset and said phase angle for said provided signal.

23. The system of claim 16, wherein said computer is further programmed to process said first portion of the carrier signal through a first linear predictor to obtain a first contribution to Doppler frequency offset for the received signal by:

computing a first correlation function

$$S1(N1;K1;T_S) = \sum_{k=1}^{N1} s1(k\cdot T_S;carr)\cdot s1((K1-k)\cdot T_S;carr)^*$$

of the provided with itself, where N1 ($\geq$1) is a first selected sample size, $K1\cdot T_S$ is a first selected time delay and $1/T_S$ is a selected signal sampling rate; and

computing a first Doppler frequency offset component

$$f_{D1} = \tan^{-1}\{S1(N1;K1;T_S)\}/(2\pi\cdot K1\cdot T_S).$$

24. The system of claim 23, wherein said computer is further programmed to process said second portion of said provided signal through a second linear predictor to obtain a second contribution to Doppler frequency offset for the received signal by:

forming a first-modified signal

$$s2(t;carr) = s1(t;carr)\ \exp(-j2\pi \cdot f_{D1} \cdot t);$$

computing a second correlation function

$$S2(N2;K2;T_S) = \sum_{k=1}^{N2} s2(k \cdot T_S;carr) \cdot s2((k-K2) \cdot T_S;carr)^*$$

of said first-modified signal with itself, where N2 (≥ 1) is a second selected sample size and $K2 \cdot T_S$ is a second selected time delay; and

computing a second Doppler frequency offset component

$$f_{D2} = \tan^{-1}\{S2(N2;K2;T_S)\}/(2\pi \cdot K2 \cdot T_S).$$

25. The system of claim 24, further comprising forming said second-modified carrier signal s3(t;carr) according to

$$s3(t;carr) = s2(t;carr)\ \exp(-j2\pi \cdot f_{D2} \cdot t).$$

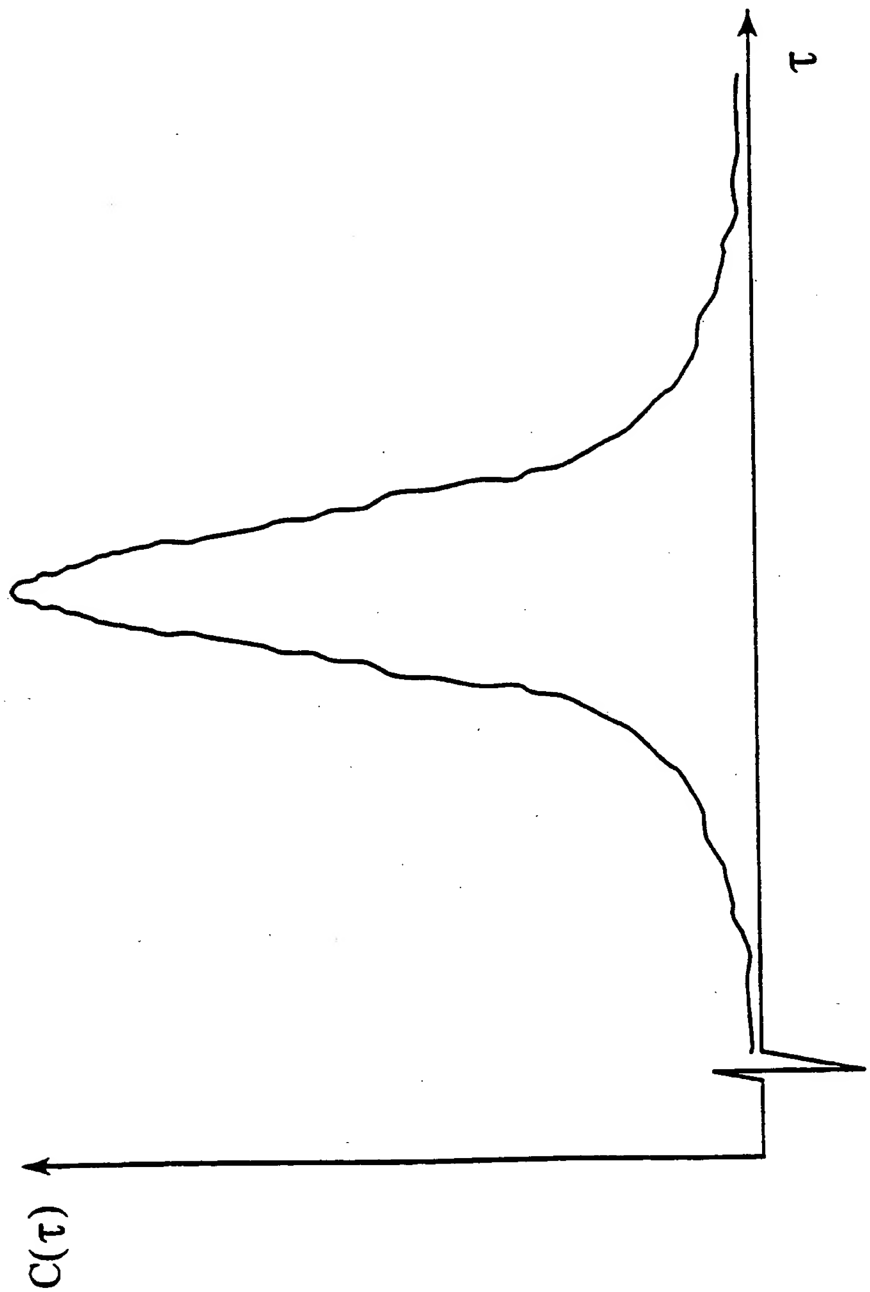

Fig. 1

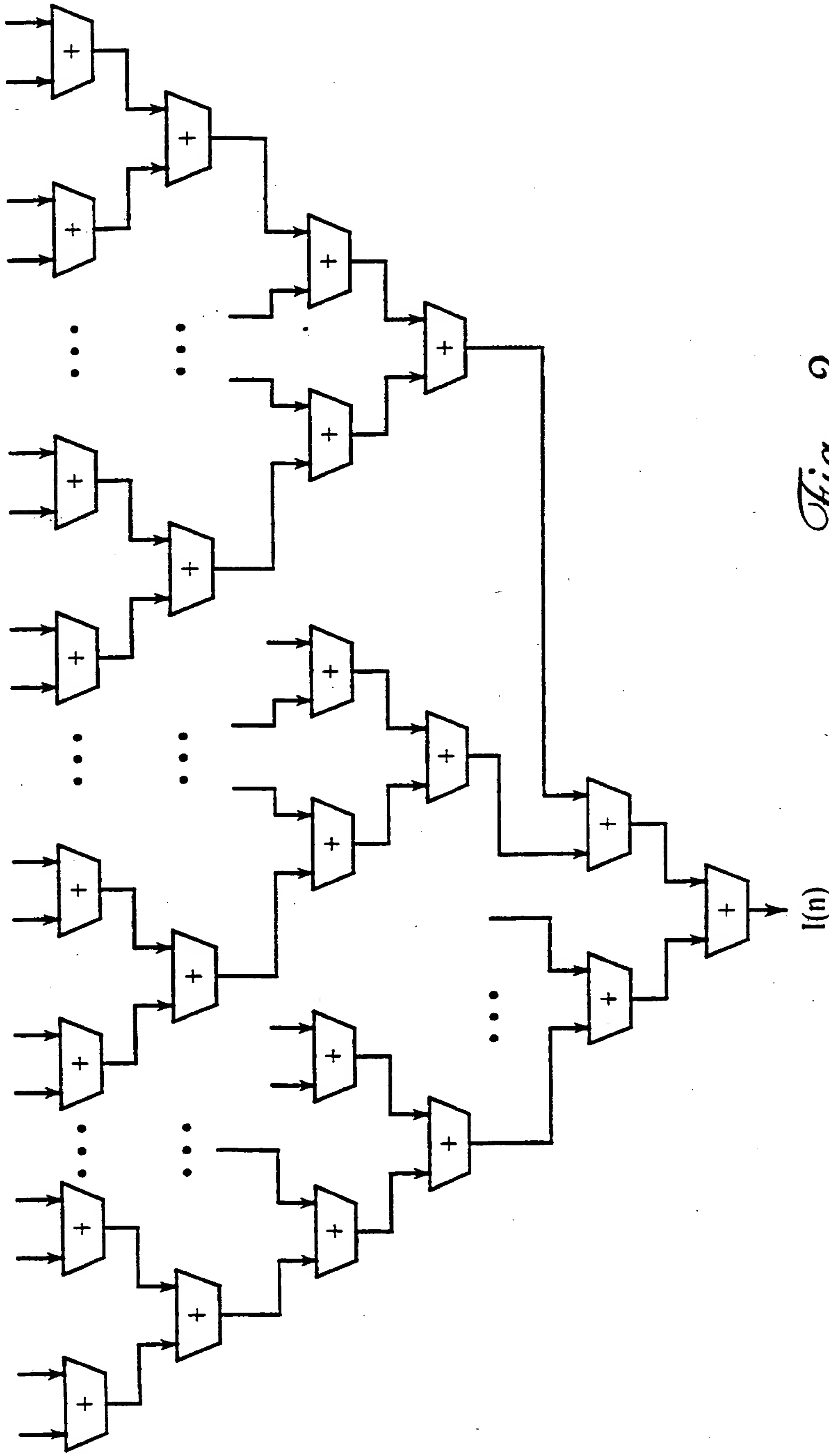

Fig. 2

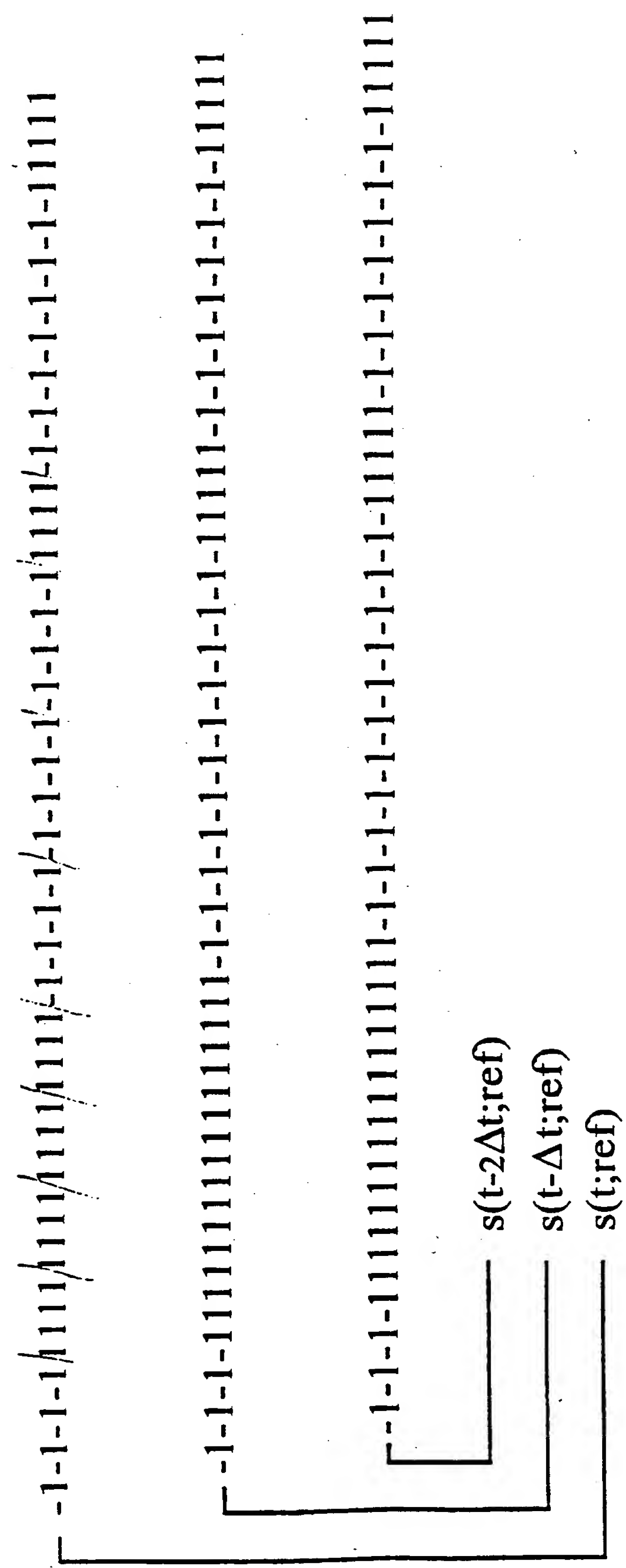

Fig. 3

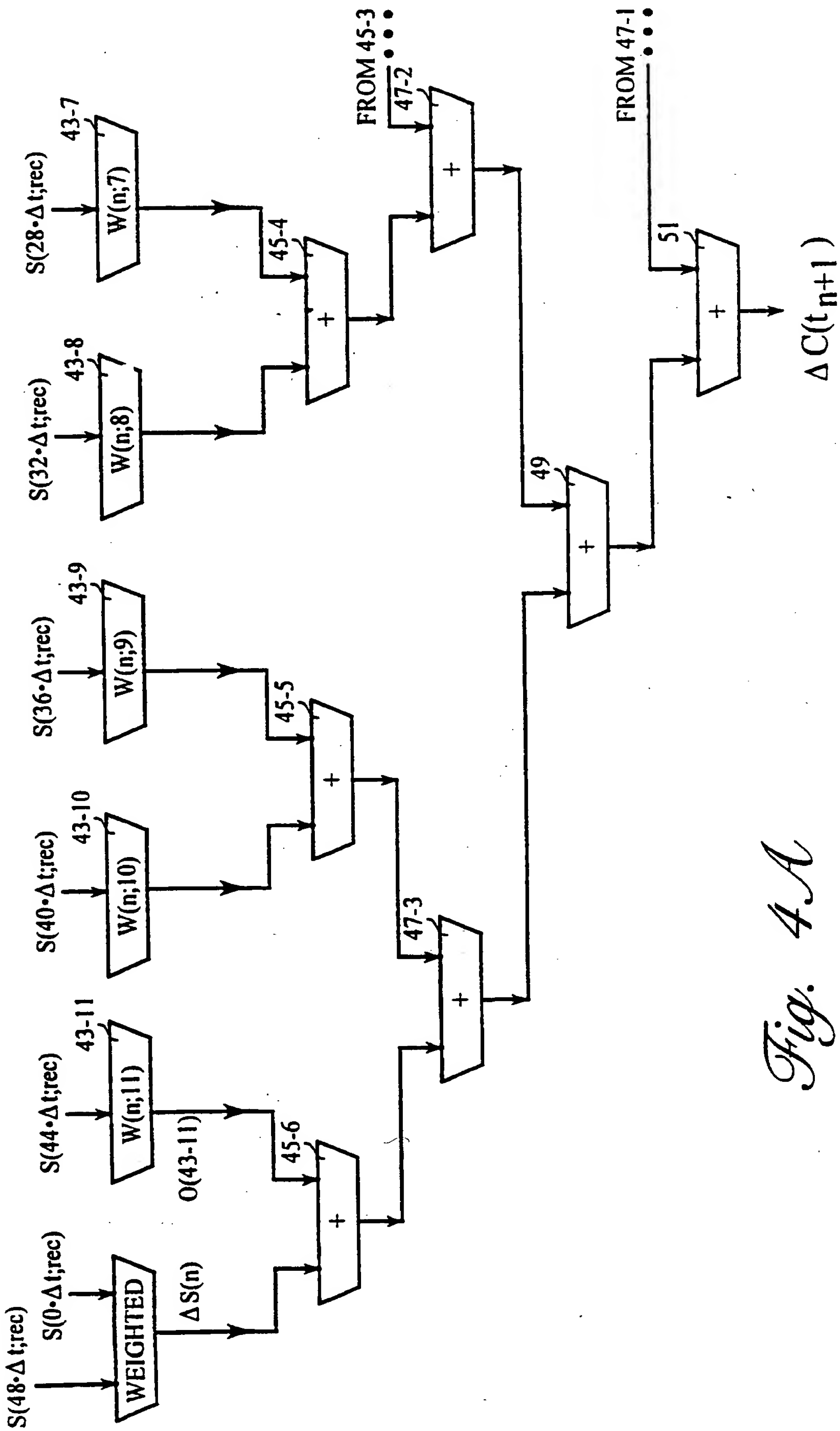
S(28•Δt;rec)
W(n;7)
43-7
S(32•Δt;rec)
W(n;8)
43-8
S(36•Δt;rec)
W(n;9)
43-9
S(40•Δt;rec)
W(n;10)
43-10
S(44•Δt;rec)
W(n;11)
43-11
O(43-11)
S(48•Δt;rec)
S(0•Δt;rec)
WEIGHTED
ΔS(n)
45-4
45-5
45-6
47-2
47-3
FROM 45-3
FROM 47-1
49
51
+
ΔC(t n+1)

Fig. 4A

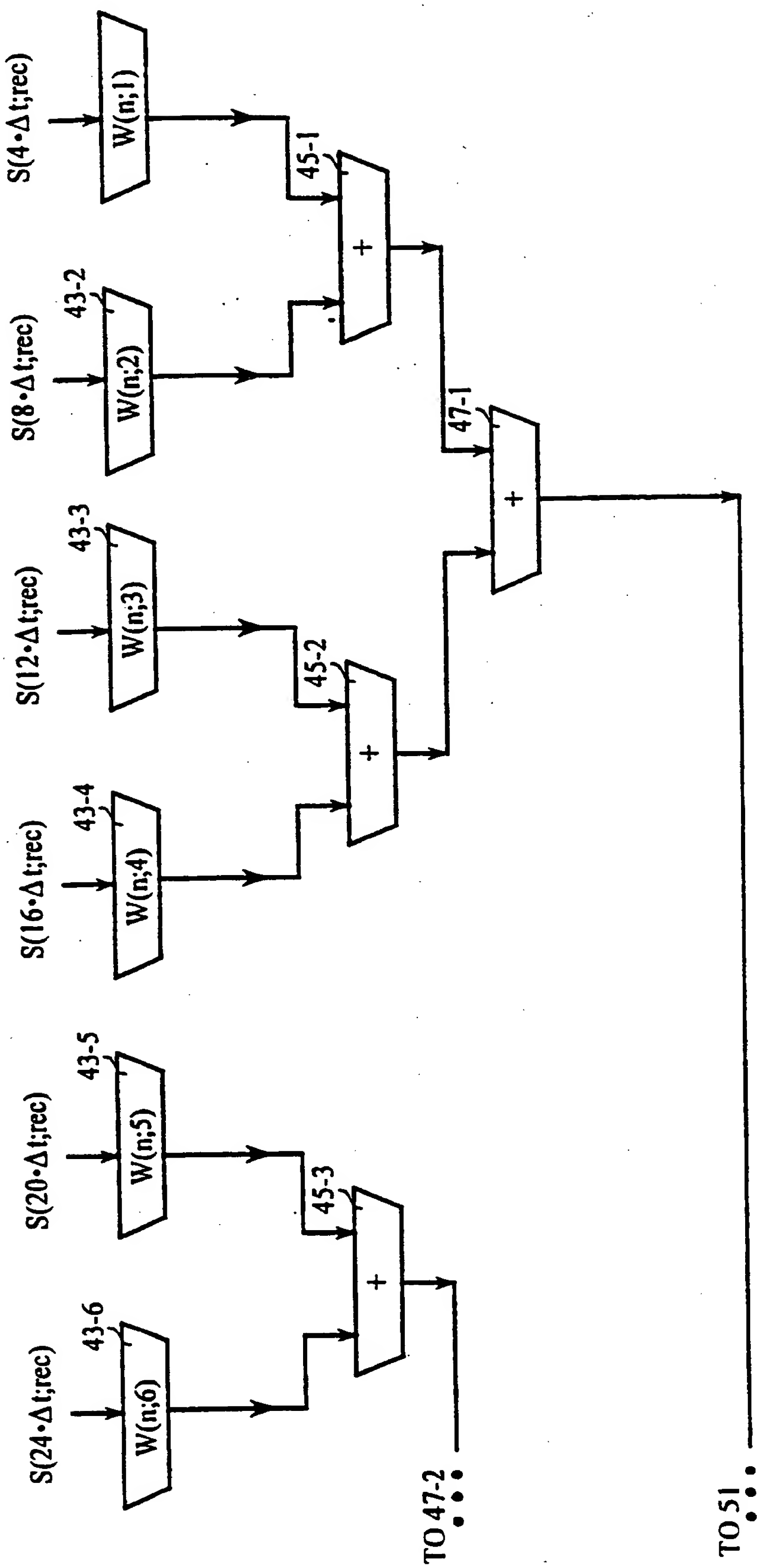
S(4•Δt;rec)
W(n;1)
S(8•Δt;rec)
43-2
W(n;2)
45-1
+
47-1
+
S(12•Δt;rec)
43-3
W(n;3)
45-2
+
S(16•Δt;rec)
43-4
W(n;4)
S(20•Δt;rec)
43-5
W(n;5)
45-3
+
S(24•Δt;rec)
43-6
W(n;6)
TO 47-2
TO 51

Fig. 4B

INTERNATIONAL SEARCH REPORT

International Application No
PCT/US 00/24823

A. CLASSIFICATION OF SUBJECT MATTER
IPC 7 H04L27/227

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)
IPC 7 H04L

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic data base consulted during the international search (name of data base and, where practical, search terms used)

WPI Data, EPO-Internal, PAJ, COMPENDEX, INSPEC

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
P,X	GB 2 344 493 A (ROKE MANOR RESEARCH) 7 June 2000 (2000-06-07) page 2, line 23 -page 3, line 21 page 8, line 16 - line 20 page 7, line 7 - line 8 page 7, line 18 - line 21 page 7, line 24 -page 8, line 7 ---	1,2,5,6, 16,17, 20,21
A	US 4 527 278 A (VAN UFFELEN JEAN-PIERRE H ET AL) 2 July 1985 (1985-07-02) page 10, column 2, line 30 - line 51 page 11, column 3, line 1 - line 11 page 14, column 10, line 23 - line 35 page 15, column 12, line 60 - line 66 page 16, column 14, line 16 - line 37 --- -/--	1,2,7-9, 11,16, 17,22,23

[X] Further documents are listed in the continuation of box C.

[X] Patent family members are listed in annex.

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"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
8 January 2001	15/01/2001

Name and mailing address of the ISA	Authorized officer
European Patent Office, P.B. 5818 Patentlaan 2 NL – 2280 HV Rijswijk Tel. (+31–70) 340–2040, Tx. 31 651 epo nl, Fax: (+31–70) 340–3016	Moreno, M

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Inter. onal Application No
PCT/US 00/24823

C.(Continuation) DOCUMENTS CONSIDERED TO BE RELEVANT

Category °	Citation of document, with indication,where appropriate, of the relevant passages	Relevant to claim No.
A	CLARK A P ET AL: "DOPPLER-SHIFT CORRECTION VIA CORRELATION FOR A RECEIVED QPSK SIGNAL" IEE PROCEEDINGS I. SOLID- STATE & ELECTRON DEVICES,, vol. 137, no. 6 PART 01, 1 December 1990 (1990-12-01), pages 387-394, XP000179891 GB,INSTITUTION OF ELECTRICAL ENGINEERS ISSN: 0956-3776 page 387, column 2, line 11 - line 20 page 388, column 2, last paragraph figure 1 -----	1,5,7, 16,20,22

2

INTERNATIONAL SEARCH REPORT

Information on patent family members

International Application No
PCT/US 00/24823

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			NO	831221	A	10-10-1983